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A Multilevel Differential Modem for Narrowband Fading Channels

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Abstract

With the recent advances in linear amplifier design, non-constant envelope modulation schemes, such as 16QAM have become a viable proposition. However, coherent carrier recovery is difficult in a mobile environment since phase lock can easily be lost when passing through a fade. One technique for overcoming this problem is to differentially encode the data in the transmitter, so that the absolute received phase is no longer important. Two receiving strategies are then possible: either a coherent receiver followed by a differential decoding process, or, more simply, a differential receiver. Since the fading is compensated without recourse to a pilot signal there are none of the delays incurred by reconstructing the channel and the receiver response time is consequently quicker. This makes differential receivers suitable candidates for use in TDM systems where a rapid response is important to utilise the allocated time slot as effectively as possible.

This paper analyses the design and performance of a differential receiver, including simulated results for operation in an existing PMR scenario in the UK. Practical considerations for the construction of such a modem are also included.

1 Introduction

Bandwidth efficient transmission of data is best achieved by using multilevel modulation schemes incorporating amplitude and phase modulation of the carrier. Until recently this has not been practical because efficient power amplifiers for the final stage of the transmitter have not been sufficiently linear and would therefore distort the signal. However, recent advances in amplifier linearisation [1] have made these signal formats a viable proposition, so schemes such as 16QAM are becoming more popular. Before 16QAM, or indeed

any other modulation scheme, can be used in a mobile environment, steps need to be taken to compensate for the amplitude and phase variations which occur in a Rayleigh fading channel. The phase change produced when passing through a fade can be extremely rapid, causing the receiver to lose lock. On reacquisition there is a strong possibility of a false lock occurring which has disastrous consequences for the system error rate.

Among the techniques used to combat this are various pilot aided schemes which are able to calibrate the receiver by correcting a known signal, be it a continuous tone as in TTIB [2] or an occasionally transmitted data symbol [3]. Another technique is to use differential encoding in the transmitter. This means that the received signal is no longer decoded as an absolute value but as a transition from one received symbol to another, and as such the absolute amplitude or phase does not matter. A coherent receiver could be used, followed by a differential decoder, or a differential receiver which does the downmixing and decoding in a single process. The latter is a much simpler receiver as there is no need for any carrier recovery, and it also has the shorter response time. Both of the pilot aided techniques require considerable signal processing, incurring significant delays, while the main delay in a differential receiver is only the duration of a single symbol. This rapid response makes a differential receiver suitable for use in TDM systems where the most effective use of the allocated time slot is desired.

The application of differential encoding to 16QAM in its traditional form (figure 1) is complicated by the fact that there are 12 unevenly spaced phases and 3 amplitudes. This produces a large number of possible phase-amplitude transitions and hence the system is very susceptible to noise. A better constellation is that shown in figure 2. Here there are only 8 evenly spaced phases and 2 amplitudes, producing 16 phase-amplitude transitions, which is much more robust in noise. Each of the 16 original symbols can therefore be easily mapped onto one of the 16 transitions.

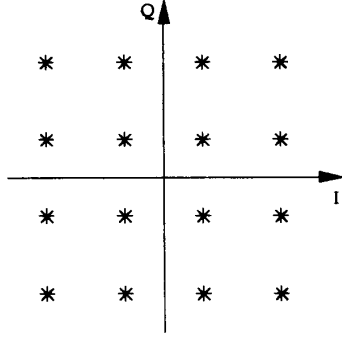


Figure 1: Traditional 16QAM Constellation

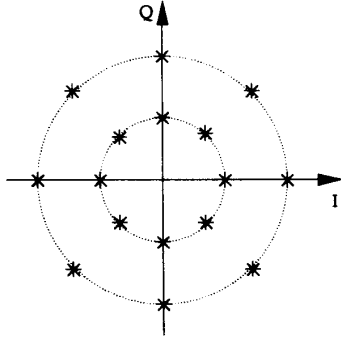


Figure 2: 16APSK Constellation

The remainder of this paper is concerned with the analysis of a modem employing differential 16APSK (or 16DAPSK). Background theory and simulation results are presented along with hardware considerations for the construction of such a modem.

2 Differential Encoding

The fundamental concept with a differential encoding scheme is that there is little change in the channel characteristics over consecutive symbols. By encoding the data as a transition from one transmitted symbol to another, the particular phase and amplitude variations imposed on the signal by the channel cease to be a problem. For example, consider a phase modulated signal

$$p(t) = \cos(2\pi f_c t + \phi(t)) \quad (1)$$

where f_c is the carrier frequency and $\phi(t)$ is the modulation phase. The channel imposes a phase distortion

$\theta(t)$ on this signal, producing

$$p(t) = \cos(2\pi f_c t + \phi(t) + \theta(t)) \quad (2)$$

If this represents the received signal for the n^{th} symbol, then the $n + 1^{\text{th}}$ symbol, a time τ later, will be

$$p(t + \tau) = \cos(2\pi f_c(t + \tau) + \phi(t + \tau) + \theta(t + \tau)) \quad (3)$$

Differential interpretation of these signals involves taking the difference between the two phase terms, $\Delta\Phi$, producing

$$\Delta\Phi = (2\pi f_c \tau + \phi(t + \tau) - \phi(t) + \theta(t + \tau) - \theta(t)) \quad (4)$$

Arranging the symbol period to be such that $f_c \tau$ is an integer, the equation reduces to a transition of the modulation phase, ϕ , with an added channel corruption, $\theta(t + \tau) - \theta(t)$. If the condition that the channel changes very little between symbols is true, then $\theta(t + \tau) \approx \theta(t)$ and the channel corruption is eliminated.

For amplitude modulation the ratio between two symbols is required since the channel corruptions are multiplicative in that dimension.

For 16DAPSK there is a combination of both amplitude and phase encoding. The phase is differentially encoded by adding the phase of the required symbol to the phase of the last transmitted symbol. The amplitude is encoded by signalling on one ring if there has been a change in amplitude and on the other if there has not. This process is explained in detail in [4].

3 Receiver Design

The basic requirement for a differential receiver is a delay block equal to the symbol period: the incoming signal is split into a direct and a delayed path, which are then multiplied together. This process brings the signal down to baseband and differentially decodes the phase modulation. For a multilevel modulation scheme like 16APSK it is convenient to work using quadrature signal components, so the differential receiver is preceded by a quadrature downconverter. This brings the signal down to an intermediate frequency, but the signal must not be brought to baseband because the differential part needs a signal which is on a carrier. Thus the basic receiver design is as shown in figure 3. The details of how this demodulates are in [4], but will be briefly repeated here.

Assume the direct signal, S , to be

$$S = d_n A_n \cos(2\pi f_i t + \phi_n + \gamma_n) \quad (5)$$

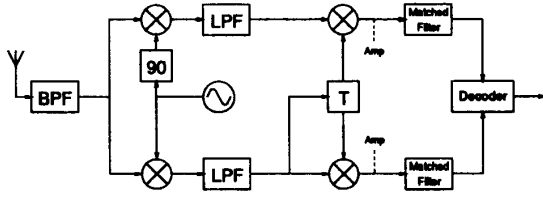


Figure 3: Basic Quadrature Differential Receiver

where A_n and ϕ_n are the amplitude and phase modulation, d_n and γ_n are the channel amplitude and phase distortion, and f_i is the intermediate frequency. The other direct signal is similar but with a *sine*-function. After delaying, mixing and matched filtering, the signals will again be *cosine* and *sine* functions, with an overall amplitude term R and a phase term Φ , where

$$R = K_m d_n d_{n-1} A_n A_{n-1}$$

$$\Phi = \phi_n - \phi_{n-1} + \gamma_n - \gamma_{n-1} \quad (6)$$

K_m represents the gain through the matched filter. Since the amplitude is the same for both signals, the phase can be determined from the arctangent of the ratio of the *sine* signal to the *cosine*. If the channel is not too severe then the phase distortion is effectively eliminated and the demodulated data, $\phi_n - \phi_{n-1}$, results. This angle is quantised to the nearest 45° and decoded into its constituent bits.

Decoding the amplitude is not so straightforward. It contains information from the n^{th} and the $n-1^{\text{th}}$ symbols, as well as the channel distortion. To remove the distortion, which is multiplicative, it needs to be divided by a previous amplitude, R_p , where

$$R_p = K_m d_{n-1} d_{n-2} A_{n-1} A_{n-2} \quad (7)$$

The division results in a ratio, Z_n :

$$Z_n = \frac{K_m d_n d_{n-1} A_n A_{n-1}}{K_m d_{n-1} d_{n-2} A_{n-1} A_{n-2}} \quad (8)$$

which reduces to

$$Z_n = \frac{d_n A_n}{d_{n-2} A_{n-2}} \quad (9)$$

If the channel does not change too rapidly, $d_n \approx d_{n-2}$ and the corruption is removed. The resulting demodulated data involves A_n and A_{n-2} , however, so the amplitude needs to be encoded over a three symbol span. Whilst this is not particularly difficult, being just a matter of software, it is not very elegant and it also means that the channel samples, d , are further apart. This reduces the degree of similarity between

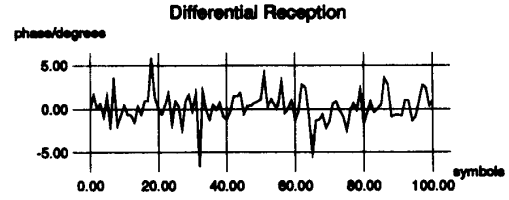


Figure 4: Phase Jitter for $\alpha = 0.9$

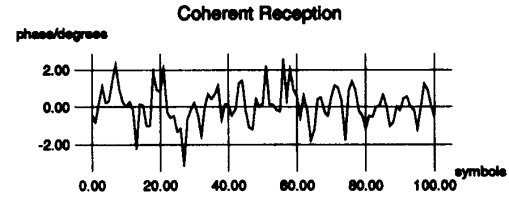


Figure 5: Phase Jitter for $\alpha = 0.9$

them which means that the fading will not be removed so effectively.

As the differential reception process multiplies the signal by itself, one symbol period delayed, it follows that the data pulse shaping used in the transmitter to improve the spectral occupancy will be squared. If a root raised cosine filter is employed in the transmitter, then for optimal noise performance an identical filter is required in the receiver, but because of the squaring the filters are no longer properly matched and the system incurs a noise penalty. The effect of this can be seen in figures 4 to 7, which show the difference between the detected and the ideal phases for a short run of random data, without any added noise. The results are shown for two roll-off factors, $\alpha = 0.9$ and $\alpha = 0.5$, where the length of the shaping function is 9 times the symbol period. Because the coherent system does not square the pulse shape and its filters are properly matched, the phase jitter is consequently smaller, although the average jitter is comparable for both cases. The amount of jitter produced for $\alpha = 0.5$ is rather more than that for $\alpha = 0.9$, a fact which could restrict the useable bandwidth efficiency.

This jitter cannot be removed by using a conventional equaliser because it is produced by a non-linear process. However, a normalising block can be inserted into the signal path to prevent the pulse shape squaring and hence produce the receiver architecture shown in figure 8. This should not affect the phase of the signal and it also eliminates the requirement for amplitude

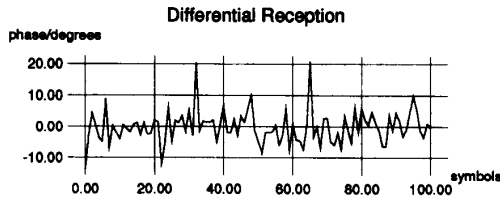


Figure 6: Phase Jitter for $\alpha = 0.5$

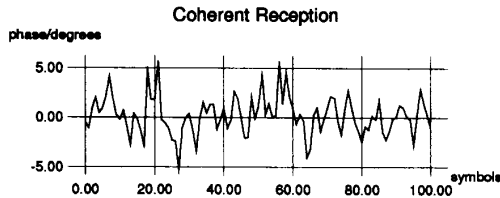


Figure 7: Phase Jitter for $\alpha = 0.5$

encoding over three symbols. The result of using this configuration with $\alpha = 0.9$ is shown in figure 9, and it is apparent that the jitter has not improved. The reason for this is that the amplitude samples which the normaliser uses are themselves imperfect, so the final result is not any better. There is therefore no benefit in using this configuration as far as the phase information is concerned and since the amplitude is derived from the same signals it will also suffer from the jitter. Consequently, this receiver design was not pursued.

An alternative point in the circuit from which to extract the amplitude information is just before the matched filters, at the points marked 'Amp' in figures 3 and 8. The signal then requires its own matched filter before the decoder. Tapping the signals in figure 3 leads to an amplitude signal requiring 3 symbol encoding and which also suffers from improper matched filtering due to pulse shape squaring. This is not desirable, but the comparable signal from figure 8 is better. Although the phase jitter is increased with this receiver design (figure 9), the amplitude signal was found to be much improved and the overall error rate was reduced.

The third way in which the amplitude can be extracted is to take the signals at the output of the quadrature downconverter. This is like having two independent receivers, one for the phase and one for the amplitude; an echo from [5] in which the 16APSK constellation is called Independent Amplitude and Phase Shift Keying. This arrangement was found to offer the

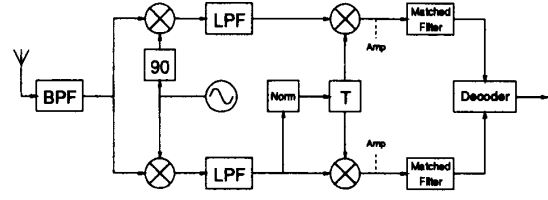


Figure 8: Basic Differential Receiver with Normaliser

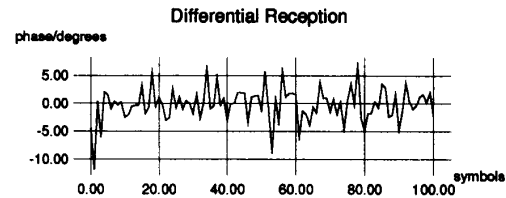


Figure 9: Phase Jitter with Normaliser for $\alpha = 0.9$

best amplitude signal, so the final receiver design was as in figure 10.

4 Simulation Results

The receiver described above was simulated in C on a Unix workstation and results taken for Additive White Gaussian Noise with and without Rayleigh fading. These are shown in figures 11 and 12. Included in figure 11 are the results for conventional 16QAM, a well documented result [6] against which comparative performance can be judged, and also results for coherent 16APSK, with and without differential encoding. For the results of figure 12, the fading was modelled by summing 8 rays with random phase shifts, amplitudes and arrival angles. The Rayleigh statistics of the resultant carrier can be measured by the simulation to check the validity of the model. The normalised Doppler value used was 1.7×10^{-3} , which corresponds to a vehicle speed of 50Km/h, a carrier frequency of 150MHz and a data rate of 16 kbits/s or 4 ksymbols/s. The carrier frequency was chosen because it matches that of an existing PMR user in the UK, and the data rate was selected as a modest, easily achievable value for preliminary investigations. The receiver timing was synchronous with the transmitter, which is a valid, if optimistic, assumption as the channel is not time dispersive. The pulse shaping used in the transmitter was a root raised cosine filter with $\alpha = 0.9$, giving a spectral efficiency of 2 bits/s/Hz and 70dB of suppression.

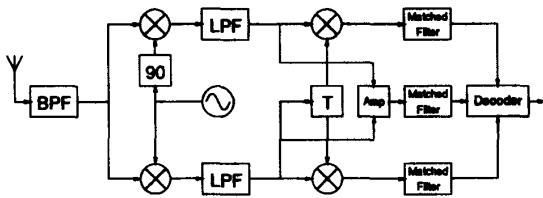


Figure 10: Receiver Architecture for 16APSK

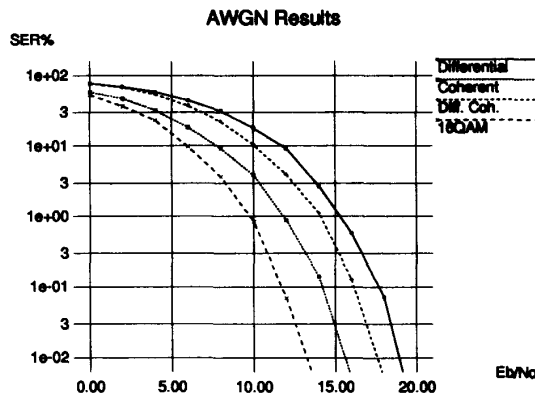


Figure 11: AWGN Results

The AWGN results without fading clearly show that coherent reception is superior in error performance to differential, with the differentially encoded case lying between the two. In fading the coherent system becomes unusable, however, with little to choose between the other two systems. On the basis of these results it would seem that differentially encoded coherent reception is the optimal technique, combining the fading performance of a differential system with a noise performance which is between 1 and 2 dB better. The carrier recovery used in the simulation is ideal, however, so in reality the differentially coherent scheme would perform worse than indicated. The trade off is thus between a noise penalty for differential reception and the added complexity of a carrier recovery scheme for the differentially coherent system.

As the normalised Doppler value increases, due to higher vehicle speeds or carrier frequencies, an irreducible error rate manifests itself. Using the same setup as before, but with a vehicle speed of 100Km/h, for example, the error rate tends towards a value of 0.2% [4]. The irreducible error rate is caused by the channel changing too quickly, with respect to the symbol period, for the differential process to remove the

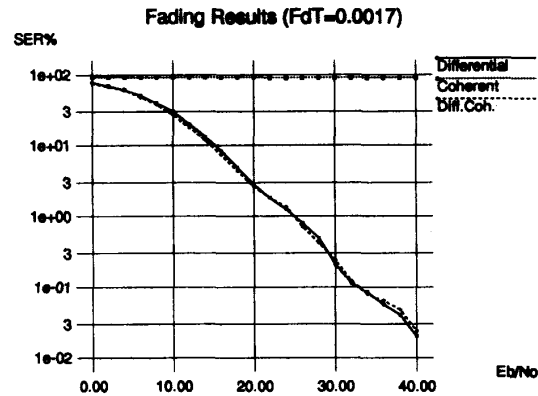


Figure 12: AWGN and Fading Results

phase and amplitude variations. Diversity has been shown in [4] to be a valuable technique for lowering the irreducible error floor, which it does by effectively slowing down the rate of change of the channel.

It was found that running the system with root raised cosine filters of $\alpha = 0.5$ produced an irreducible error rate due to the imperfect matching and consequent jitter mentioned in the previous section. If the value of α is increased to 0.7 then the residual error disappears. This means that the system must operate with $\alpha \geq 0.7$, with a maximum bandwidth efficiency of approximately 2.3 bits/s/Hz.

5 Hardware Considerations

A modem has been partially constructed following the design in figure 10, using two Texas Instruments TMS 320C25 processors: one in the transmitter and one in the receiver. The transmitter uses a 22.5° rotated version of the constellation in figure 2, but apart from that it operates exactly as described above and in [4]. There is no rf stage at present, the transmitted signal being output on a low frequency DSP generated carrier. The receiver takes this signal and a quadrature version of the same as its inputs, assuming them to have come from the quadrature downconverter. The differential stage then follows, along with the matched filters and decoding of the received symbols back into a bit stream.

Because the input to the receiver is not a baseband signal, and cannot be for the differential part to function correctly, the receiver must operate at a higher sampling frequency than is strictly necessary for the

given data rate. This restricts the amount of processing bandwidth available to the processor and hence the maximum data rate which could be supported. Implementing the differential stage as an analogue process would remedy this situation, but the symbol period delay required could be difficult to achieve. There are SAW filters available with a group delay of the required order of magnitude [7], but the effect of using these has not been established. It is important that the delay is exactly one symbol period for the demodulation process to work properly, something which is easy to achieve digitally as it is simply a matter of a shift register of the required length.

The other main use of processing time is the matched filtering. To have a well confined power spectrum the time domain response of the shaping filter needs to be several times the symbol period. (In the simulation a filter length of 9 times the symbol period produced 70dB rejection.) This means that for proper matched filtering the receiver filters must be the same length, and using the receiver architecture in figure 10 there have to be three of them. It is possible that a separate processor could be used as an accelerator for the matched filters (eg the Inmos A100), which would then relieve the main processor for its equally important scheduling and decoding duties.

A useful feature of the separate amplitude receiver design is that in the calculation of the amplitude signal, by squaring and summing the quadrature downconverter outputs, a good basis for timing recovery is also provided. The square of the amplitude signal contains a spectral line at the symbol rate, which when isolated can be used to drive the sample clock. This does require some stringent filtering, which again might be better done as an analogue process if processing bandwidth is tight. The current implementation uses a moderate bandpass filter followed by a phase locked loop, all implemented in the DSP.

6 Conclusions

An analysis of differential receivers has been presented, leading to the conclusion that, for the scenario indicated, the best receiver architecture is as shown in figure 10 where there is a separate amplitude extraction circuit. Results produced by simulation have been presented showing the performance of this receiver in AWGN and also in Rayleigh fading.

On the practical side, using a separate amplitude extraction circuit does mean that extra stages are required, but the basis of a timing recovery signal is produced at the same time. As a timing recovery system

would be required anyway, the extra stages are quite justified. By doing as much of the processing as possible using analogue electronics the available processing bandwidth can be increased, thereby allowing higher data rates to be supported by the available technology. Using accelerators to do the matched filtering would further enhance performance.

7 Acknowledgements

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